DESIGN AND FABRICATION OF THRESHOLD EXTENSION DEMODULATOR FOR 70 MHz I. F. F. D. M.—F. M. SYSTEMS

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C E R T I F I C A T E

Certified that this work "Design and Fabrication of Threshold Extension Demodulator for 70 MHz IF FDM-FM Systems", by Mr. K.R. Srivathsan has been carried out under my supervision and that this hasnot been submitted elsewhere for a degree.

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SYNOPSIS

An effort has been made to design and fabricate a Threshold Extension Demodulator (TED) for wideband FDM-FM Radio Links accomodating upto 120 voice channels. The use of TEDs instead of conventional demodulators for FM results in a substantial improvement of receiver threshold and an increase in the dynamic range of input carrier-to-noise ratio over which the advantages of FM are available.

The basic theory of FM receivers, the phenomenon of FM threshold, the philosophy behind FM threshold extension and the various basic configurations of TEDs are discussed in Chapter-I. The principle of operation, the fundamental design equations, and the basic configuration of the Phase Lock Loop (PLL) as a TED are explained in Chapter-II.

In Chapter-III, the design of PLL demodulator is carried out. It is shown that for the system under consideration a threshold improvement of 6-7 dB is possible. The design and implementation of individual loop components, i.e. phase detector, voltage controlled oscillator (VCO), loop filter and amplifier, post-detection filter and other auxiliary parts, are discussed. The advantages of a push-pull VCO and its implementation are described in detail. The elements of operation and testing are presented.

C O N T E N T S

CHAPTER I — Page No.

1.1 Introduction 1

Frequency Modulation

Conventional Demodulation Techniques

1.2 Reception under Noise 3

Representation of Signal with Noise

Threshold Effects

1.3 Threshold Extension Demodulators 12

1.4 Other Threshold Extension Techniques 15

CHAPTER II

2.1 Principle of Operation 17

Introduction

Non-linear Operating Conditions in the PLL

2.2 Design of PLD for Threshold Extension 23

Minimum Loop Noise Bandwidth

Threshold Conditions in PLD

Post Detection Filtering

Distortions in the PLD

Delay Compensation

Conclusion

CHAPTER III   Page No.

3.1 Design of PLD for a 120 Voice Channel

FDM-FM System 33

System Design

PLD Design

3.2 Design of Loop Components 35

Voltage Controlled Oscillator

Push-Pull VCO

Baseband input and varactor bias Amplifier

Phase Detector

Loop Filter and Amplifier

Post-detection filter

3.3 Closing the Loop 51

3.4 Testing the PLD 51

APPENDIX 58

CHAPTER I

1.1 INTRODUCTION

Frequency Modulation:

The earliest and one of the most widely used analog modulation schemes which has the property of exchanging transmission bandwidth to demodulated signal-to-noise Ratio (SNR) is Frequency Modulation (FM).

In FM, the modulated signal is of the form:

$$e(t) = A \cos \left(w_c t + \Delta w_p \int_0^t f(t)\, dt\right) \qquad (1.1)$$

where, w_c = Carrier frequency in rad./sec.

$f(t)$ = Information or Transmitted Baseband signal; we assume that $|f(t)| \leq 1$.

Δw_p = Maximum frequency deviation in rad./sec. of the carrier from its centre frequency w_c.

This is a special case of the general exponential modulation schemes. If we interpret instantaneous frequency as the rate of change of phase of the signal, for FM, we have

Instantaneous Phase: $\phi_i(t) = \Delta w_p \int_0^t f(t) dt \qquad (1.2a)$

Instantaneous Frequency: $w_i = w_c + \dfrac{d\phi_i(t)}{dt}$

$$= w_c + \Delta w_p f(t) \qquad (1.2b)$$

The Fourier Spectrum of e(t) given the spectrum of f(t)[1] is rather complicated to arrive at. However, if f(t) is low pass in nature and band limited to 'w_b' rad./sec. then, e(t) is band pass, centred around 'w_c' in nature and a good approximation to the Intermediate Frequency (IF) Bandwidth of e(t) is given by the Carson's Rule:

$$B_T = 2\,(f_b + \Delta f_p) = 2f_b(1+m_p). \qquad (1.3)$$

where B_T = IF Bandwidth of the modulated signal in Hertz.

$f_b = w_b/2\pi$, top baseband frequency in Hertz.

$f_p = \Delta w_p/2\pi$, peak frequency deviation in Hertz.

$m_p = \Delta w_p/w_b$, peak modulation index of e(t).

Conventional Demodulation Techniques:

All conventional methods to extract f(t) from e(t) as given in (1.1) can be described mathematically to consist of differentiation followed by envelope detection. This is readily seen by differentiating e(t):

$$\dot{e}(t) = -A(w_c + \Delta w_p f(t)\ \mathrm{Sin}(w_c t + \Delta w_p \int_0^t f(t)dt) \qquad (1.4)$$

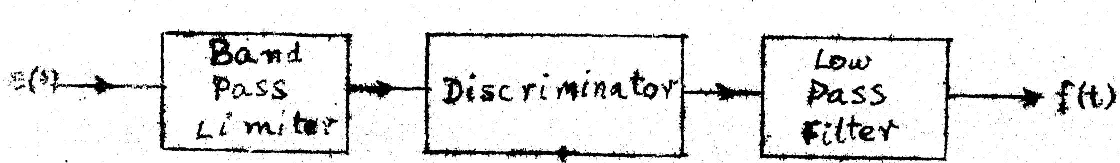


Fig. 1.1

Two practical methods using this principle are shown in Figs. 1.1 and 1.2.

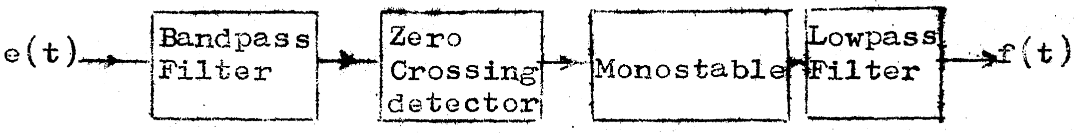


Fig. 1.2

The Bandpass Limiter in Fig. 1.1 has a bandwidth B_T centred around w_c and limits any amplitude variation in e(t). The Linear discriminator (LD) is usually an offset double-tuned circuit whose outputs are peak detected and summed to balance out the DC term due to w_c. (Fig. 1.3)

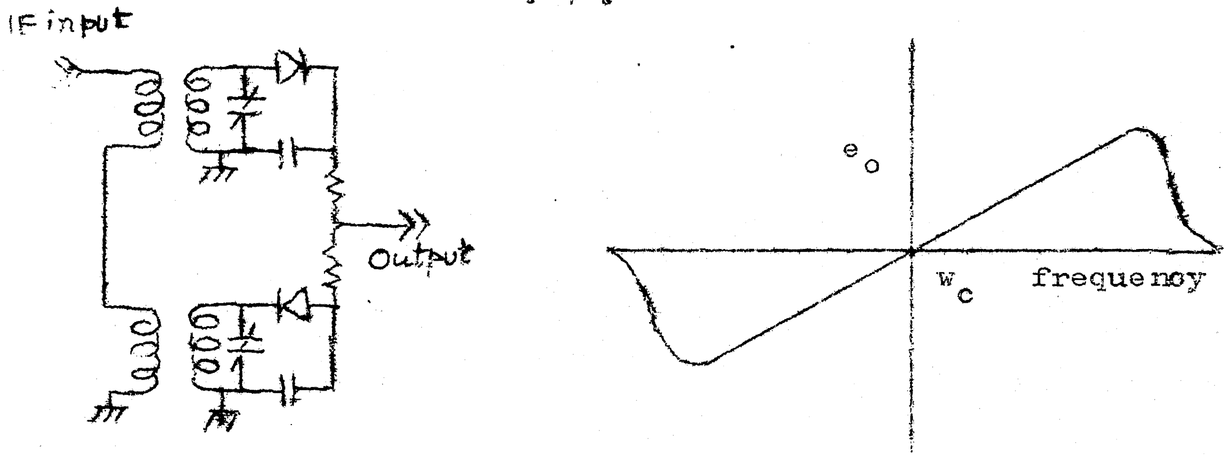
1.2 Reception Under Noise:

Representation of signal with noise[1]:

Usually, the input to the FM demodulator, is not the simple function e(t) as given in 1.1 alone, but is of the form,

$$r(t) = e(t) + N(t) \qquad (1.5)$$

where N(t) is an additive random noise. N(t) is, in most practical cases assumed to be zero mean, wide sense stationary, white, Gaussian noise of double sided spectral density $N_o/2$ watts/Hz. and bandlimited by the Bandpass Filter to



(a) Discriminator (b) Transfer characteristic

Fig. 1.3

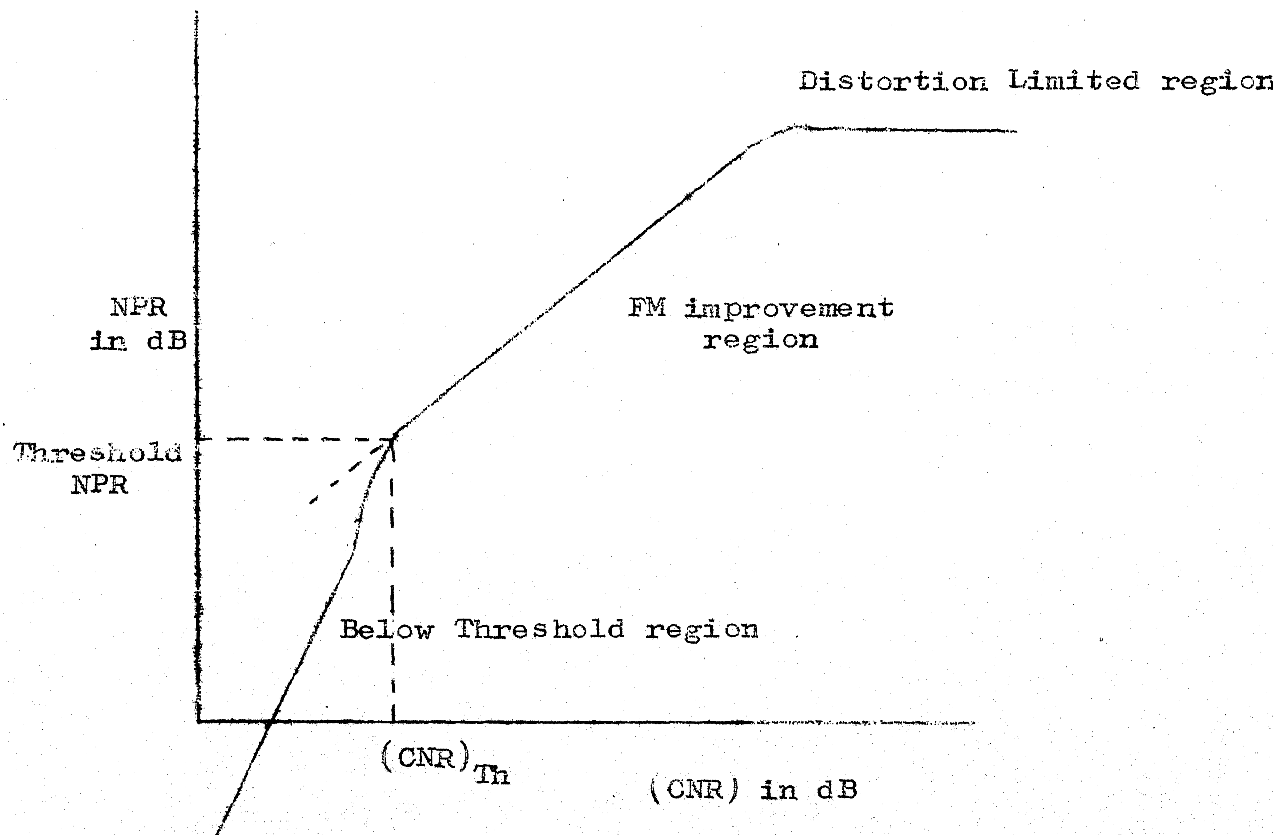


Fig. 1.6

the IF Bandwidth B_T.

Since $N(t)$ is a narrowband noise, we may write,

$$N(t) = n_c(t) \operatorname{Cos} w_c t - n_s(t) \operatorname{Sin} w_c t \qquad (1.6)$$

where $n_c(t)$ and $n_s(t)$ are Gaussian and Lowpass with Spectral height N_o watts/Hz. over the band $-B_T/2$ to $B_T/2$ Hertz. Using (1.6) we can rewrite (1.5) as:

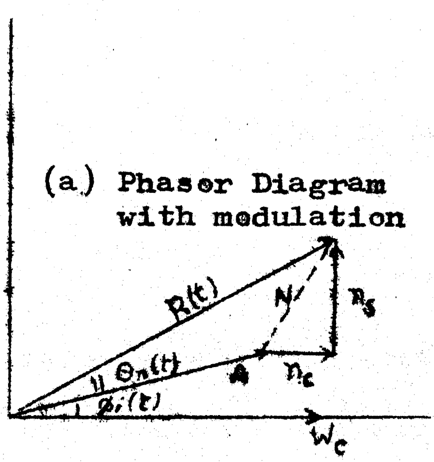
$$r(t) = R(t) \operatorname{Cos}(w_c t + \phi_i(t) + \theta_n(t)) \qquad (1.7)$$

where, $$\theta_n(t) = \operatorname{Tan}^{-1} \frac{n_s(t) \operatorname{Cos} \phi_i(t) - n_c(t)\operatorname{Sin}\phi_i(t)}{A + n_c(t) \operatorname{Cos} \phi_i(t) + n_s(t)\operatorname{Sin}\phi_i(t)} \qquad (1.8)$$

$\theta_n(t)$ = is the phase modulation due to noise.

$R(t)$ is the modulation in the envelope of $e(t)$ due to noise.

The above relations are derived directly from the Phasor representation of $r(t)$ in Fig. 1.4(a). In all FM demodulators



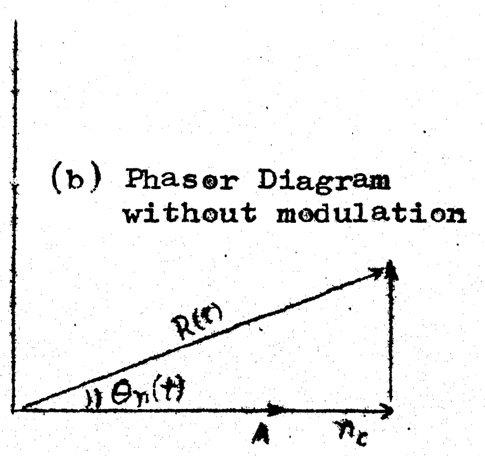


Fig. 1.4

R(t) is taken care by limiting. Only the phase noise $\theta_n(t)$ is detected and passed on to the baseband. In order to simplify analysis, we assume an unmodulated carrier. (Fig. 1.4(b)), i.e.

$$\phi_i(t) = 0$$

$$\theta_n(t) = \tan^{-1} \frac{n_s(t)}{A + n_c(t)} \tag{1.9}$$

High CNR input condition:

In this case, $A \gg |n_c(t)|$. Then $\theta_n(t)$ may be approximated as,

$$\theta_n(t) = \tan^{-1} \frac{n_s(t)}{A} = \frac{n_s(t)}{A}$$

The detected noise is $\dot{\theta}_n(t) = \dfrac{\dot{n}_s(t)}{A}$ (1.10)

The output noise spectrum is, therefore, parabolic, given by,

$$S_o(f) = \frac{f^2}{A^2} \; N_o \text{ watts/Hz} \tag{1.11}$$

For a simple baseband signal with top baseband frequency of 'f_b' Cps., total detected noise power, (as shown in Fig. 1.5)

$$P_N = \frac{N_o}{A^2} \int_{-f_b}^{f_b} f^2 df = (2/3) \frac{N_o}{A^2} f_b^3 \text{ watts} \tag{1.12}$$

For a single tone modulation at 'f_b' Cps, with index m_p, the output signal power is:

$$P_s = \frac{m_p^2 f_b^2}{2} \quad \text{watts}$$

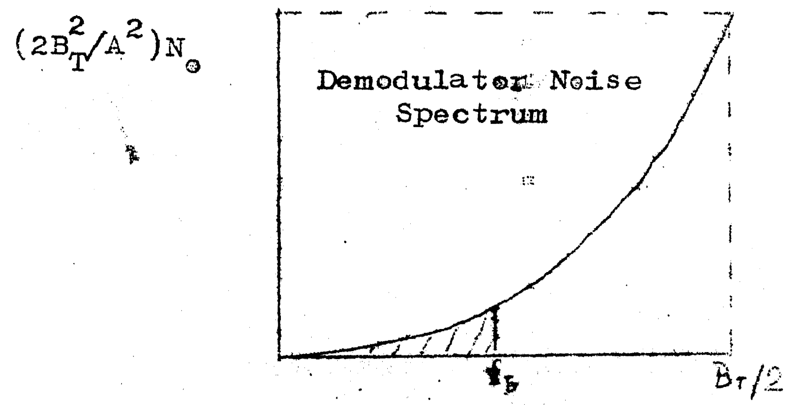


Fig. 1.5

Hence, output SNR for test tone modulation is

$$(SNR)_{TT} = P_s/P_N = 3\ m_p^{\ 2}\ (CNR)_{AM} \qquad (1.13)$$

where $$(CNR)_{AM} = \frac{A^2/2}{2N_o f_b} \qquad (1.14)$$

is the input carrier to noise ratio (CNR) over an equivalent Amplitude Modulation (AM) bandwidth. It is obvious from (1.13) that for $m_p > 1$ FM gives better SNR than AM.

For Frequency Division Multiplex (FDM) signals, consisting of linearly added spectrally non-overlapped channels, the output signals noise ratio in any given channel is measured in terms of Noise Power Ratio (NPR) defined as the ratio of output noise power in the given channel when the transmitted signal is modulated with that channel loaded alongwith other channels, to the output noise power when

the loading in that channel is removed ahead of transmission by a bandstop filter. Also since FDM signals are noise-like, they are characterised by R.M.S. deviation and R.M.S. modulation index 'σ', defined as

$$\sigma = \frac{\text{RMS frequency deviation}}{\text{Top Baseband frequency}} = \left(\frac{\Delta w_{rms}}{w_b}\right)$$

In a high 'Q' FDM channel ($f_{CH} \gg$ channel Bandwidth) centred at the freqnency 'f_{CH}',

One sided power spectral density of noise (from 1.11)

$$= 2 \frac{f_{CH}^2}{A^2} N_o$$

The power spectral density of signal $= \dfrac{(\Delta f_{rms})^2}{(f_b - f_a)}$

Taking the ratio of the two, we get

$$NPR = \frac{2\sigma^2}{(1-f_a/f_b)} (fb/f_{CH})^2 (CNR)_{AM} \qquad (1.15)$$

where f_a is the bottom Baseband Frequency.

The relation described in 1.15 is shown in Fig. 1.6 with the title 'EM improvement region'. Specifications[2] for FDM signals usually prescribe minimum NPR for the worst channel, which is usually the top most channel due to the parabolic FM noise spectrum. All practical demodulators have some amount of distortion due to phase nonlinearities in

the bandpass response and amplitude nonlinearities in the various stages. This results in self generated noise due to intermodulation and harmonic distortion of the baseband FDM signal. This predominates over the additive channel noise at high CNR as shown by the saturation in the 'Distortion Limited Region' of Fig. 1.6.

Threshold Effects[3,4]:

In all FM demodulators, below a certain threshold value of input CNR, which depends on the type of demodulation scheme, the output SNR deteriorates much faster with fall in input CNR than those predicted by the linear relations(1.13) and (1.14). For the purpose of measurement, the threshold CNR may be defined[5] as the value at which the output SNR (or NPR for FDM signals) falls 1 dB below that predicted by the relations (1.13) and (1.14).

The onset of threshold arises due to the invalidity of the approximation in (1.9) under low CNR conditions and is punctuated by the occurrence of some complex phenomena. An exact analysis should be able to consider the spectrum of the derivative of $\theta_n(t)$ as given in (1.7). However, this is mathematically complicated. Much work[4] has been done to predict the relations in the threshold region. One widely used analysis given by S.O. Rice[3], and which predicts

reasonably accurate values for threshold is briefly described below:

Consider the Phasor diagram of an unmodulated carrier with noise (Fig. 1.7). At the demodulator input, the tip of the resultant vector 'P' wanders randomly about the carrier tip 'A'. As the CNR decreases, the phase noise $\theta_n(t)$ goes through larger excursions. This increases the rate at which the locus of the point 'P' encircles the origin in a given period of time. For every encirclement of the origin, $\theta_n(t)$, goes through a step transition of $\pm 2\pi$ and this produces impulses called 'clicks' or 'Threshold Impulses' (ThI) in $\dot{\theta}_n(t)$ of strength 2π at the output of a discriminator. These impulses contribute a large flat low pass spectrum which is added on to the parabolic noise spectrum due to the locus of P non-encircling the origin O (Fig. 1.8). If the locus of P does not encircle the origin but goes through large transitions, then $\theta_n(t)$ essantially consists of impulses or doublets with a maximum height of $\pm\pi$ rads., the derivative spectrum of which is close to parabolic. Hence, they do not contribute much baseband noise when compared to the encirclement noise. Rice[3] has shown that the expected number of ThI per second for an unmodulated carrier is

$$N = r \,\mathrm{Erfc}(\sqrt{p}) \qquad (1.16)$$

where $$r^2 = \frac{\int_0^{\infty}(f-f_c)^2 \, S_i(f)df}{\int_0^{\infty} S_i(f)\, df} \qquad (1.17)$$

is the radius of gyration of the input noise spectrum

$$p = \frac{A^2}{2N^2(t)} \text{ is the CNR at limiter input} \qquad (1.18)$$

$$\mathrm{Erfc}(x) = \frac{2}{\sqrt{\pi}} \int_x^{\infty} e^{-a^2} da$$

Since each impulse of strength '2π' has a double sided energy spectrum of $(2\pi)^2$, the power spectral density of ThI is $N\,(2\pi)^2$. Hence, total noise power due to both ThI and the regular jitter of $\theta_n(t)$ is, to a good approximation given by, adding (1.11) and the term $(8\pi^2 Nf_b)$:

$$P_N' = (2/3)\,\frac{N_o}{A^2}\, f_b^3 + 8\pi^2 Nf_b \qquad (1.19)$$

This changes, the expression for Test tone SNR in (1.12) as:

$$(SNR)_{TT} = \frac{(\Delta f_p)^2/2}{(f_b^2/6(CNR)_{AM}) + 8\pi^2 Nf_b} \qquad (1.20)$$

With modulation, the ThI rate increases slightly, and both (1.13) and (1.18) are still very good approximations. Crosby[5] has shown that, with modulation, the 'clicks' occur mainly in the direction opposing the signal. Frankle[7] has shown that the ThI rate increases by an equivalent of about 0.5 dB

of received carrier power under noise modulation. The threshold and below threshold region are shown in Fig. 1.5. When received carrier power is less than the noise power, the carrier and noise interchange their roles in the Phasor diagram of Fig. 1.4. Under this condition, called 'Capture Effect'[1], the demodulator output consists mainly of noise.

1.3 Threshold Extension Demodulators[8]:

Conventional demodulators fail to utilize the important a priori information in an FM signal, i.e. although the carrier frequency may vary over a wideband, its rate of change can not be greater than the baseband rate. This fact exposes the input of the discriminator to the noise energy in the entire IF bandwidth, thereby increasing the probability of occurrence ThI even at relatively high CNR levels. By using some kind of 'adaptive' or 'tracking filters' which respond to frequency changes only around the instantaneous frequency at the baseband rate, it must be possible to push down the threshold CNR value in large index systems.

In practice, there are three basic configurations for the threshold Extension Demodulator (TED). They are shown in Fig. 1.9.

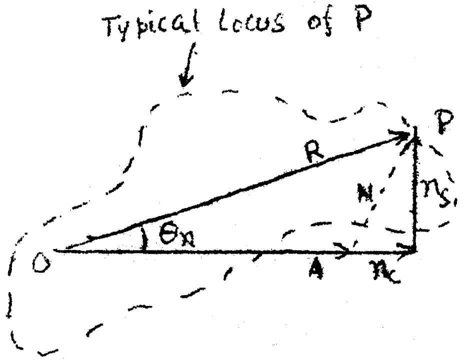


Fig - 1.7

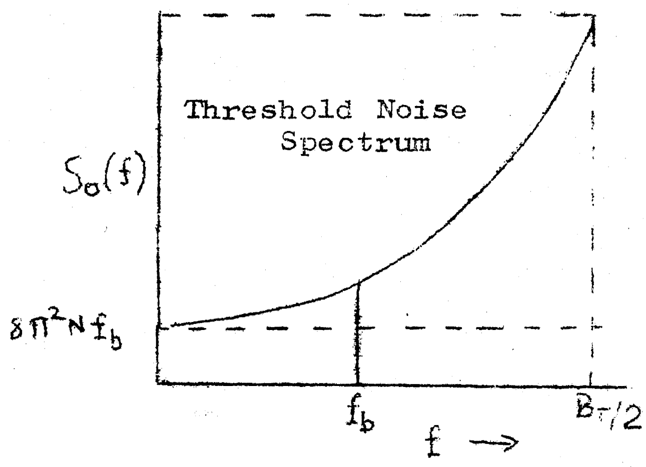
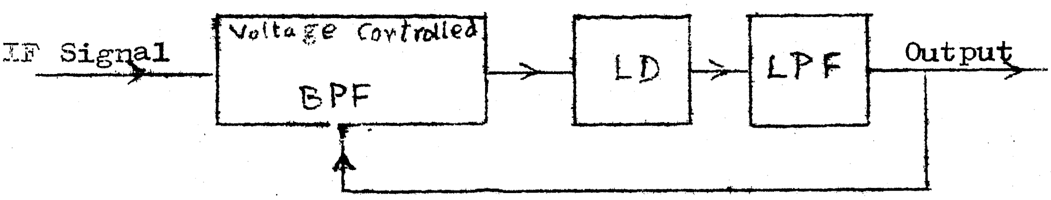
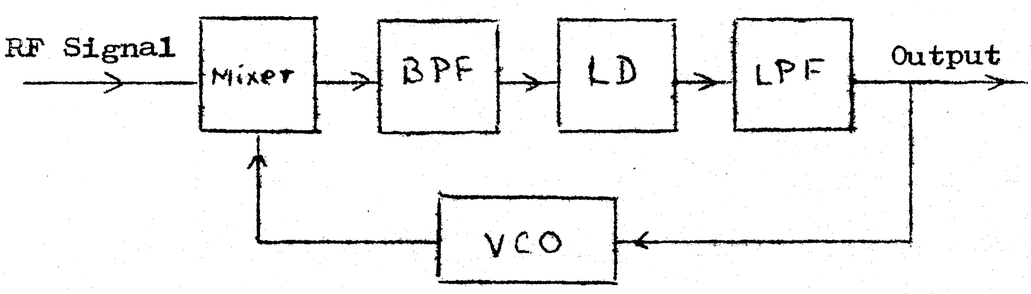


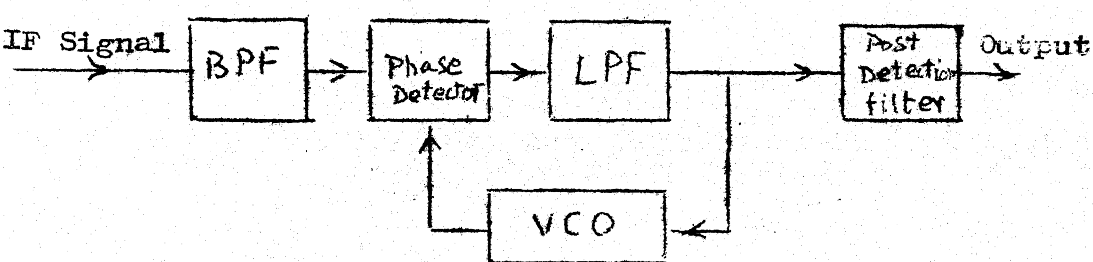
Fig - 1.8



(a) Tracking Filter Detector



(b) Frequency Modulation Feedback Detector



(c) Phase Lock Demodulator

Fig. 1.9

In the 'Tracking Filter Demodulator' (Fig. 1.9a), the demodulated signal output is fedback to control the centre frequency of a narrow band voltage controlled filter which preceedes the usual Limiter-Discriminator. The main difficulty is the construction of a linear voltage controlled band-pass filter. In the 'Frequency Modulation Feedback Demodulator' (FMFB, Fig. 1.9b), the wideband FM is compressed to a narrowband FM at the IF by modulating voltage controlled local oscillator (VCO) by the detected signal output. The narrowband FM is demodulated through a narrowband conventional demodulator. In this case, the design of the IF system is integrated with the design of Local Oscillator and RF mixer stages. It has been found that for all types of FM demodulaters using feedback, the loop delay should be as low as possible. This condition makes the design of loop components for FMFB a little complicated due to more number of critical stages within the feedback loop. However, there are more number of commercially successful demodulators of this and the Phase Lock types used in wide-band Radio Links and Satellite Communications than those using the 'Tracking Filter' principle.

In the Phase Lock Demodulator (PLD), the VCO is locked to the IF by an Automatic Phase Control Loop and

the VCO faithfully tracks the carrier at all instants. Then the control input voltage to the VCO is proportional to the deviation of the carrier and hence becomes the demodulated output. In practice, well designed demodulators of all the three types have been found to give same order of threshold extension in the range 3 to 10 dB, depending upon the modulation index and a variety of other factors.

1.4 Other Threshold Extension Techniques:

Studies[8] have indicated that in practical threshold extenders using the above mentioned schemes, factors like loop delay, phase nonlinearities in the IF filters, non-linear characteristics of VCO and phase detectors, adversely affect the threshold improvement possible. Attempts to improve the threshold by shaping the PD characteristics in PLD have led to 'Tanlock' and Extended Range PLDs. It has been found that TEDs with better threshold can be built by multiple FMFB loops or compound loops like FMFB-ERPLD and similar ones.

Efforts have also gone in to reduce the severity of noise spikes[9] which occur in FM demodulator output by utilizing non-linear filters like Delta-modulators or sampling techniques. The non-linear filters selectively isolate the severity of 'Clicks' while allowing the signal to pass through. Some reports[9] claim as much as 3 dB improvement by such techniques.

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CHAPTER II

DESIGN OF PLD FOR LOW THRESHOLD DEMODULATION

In this chapter, the basic theory of operation and design of PLD for low threshold and high saturation NPR are presented.

2.1 Principle of Operation:

Introduction:

The basic PLD consists of a (i) Low Pass Filter a (ii) Phase Detector and a (iii) Voltage Controlled Oscillator (Fig. 2.1). The functions at the inputs, and output of the PD and the output of the LPF are shown in Fig. 2.1.

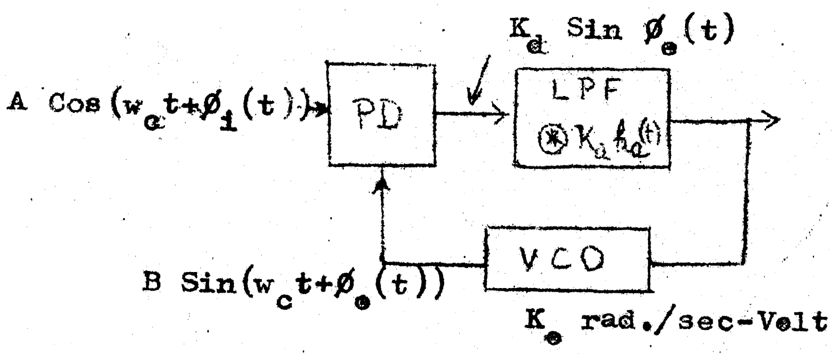


Fig. 2.1.

Most of the practical phase detectors are either the multiplier or the sampling types. In such cases, the PD relation is given by (Fig. 2.2.)

$$b(t) = K_d \ Sin \ (\phi_i(t) - \phi_o(t)) = K_d \ Sin \ \phi_e(t) \quad (2.1)$$

where $\phi_i(t)$ = phase function at signal input

$\phi_o(t)$ = phase function at reference input

$\phi_e(t) = \phi_i(t) - \phi_o(t)$ phase error between signal and reference.

This assumes that there is a static phase error of $\pi/2$ rads. between signal and reference.

The frequency deviation of the VCO is given by:

$$\frac{d\phi_o(t)}{dt} = K_o \; C(t) \tag{2.2}$$

Applying Laplace Transformation, we have

$$\phi_o(s) = K_o \; C(s)/s \tag{2.3}$$

If we assume that the phase error is small, i.e. if $|\phi_e(t)| \ll \pi/2$ we have the linear approximation for (2.1)

$$b(t) = K_d \; \phi_e(t) \tag{2.4}$$

Linear Model:

The Linear Model has the equivalent circuit shown in Fig. 2.3.

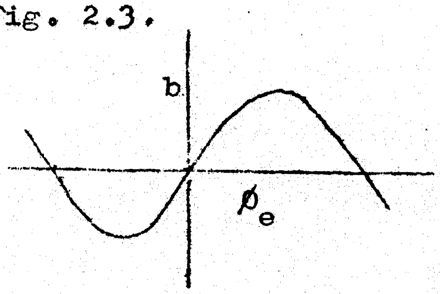


Fig. 2.2

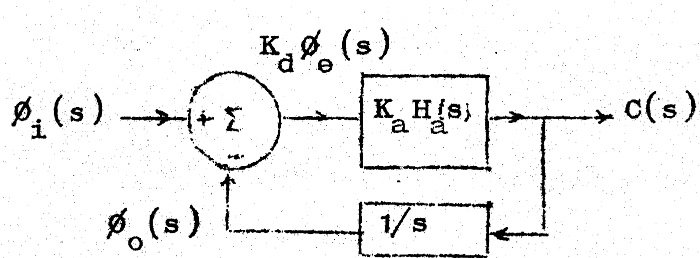


Fig. 2.3

This model holds iff. $|\phi_e(t)| \ll \pi/2$ rads. Otherwise, the nonlinear and periodic nature of the PL gives the following differential equation for the Loop:

$$\frac{d\phi_o(t)}{dt} = K_a K_d K_o \; h_a(t) \circledast \mathrm{Sin}\; (\phi_i(t) - \phi_o(t)) \qquad (2.5)$$

where $h_a(t)$ is the inverse transform of $H_a(s)$ and $\circledast$ denotes the convolution operator.

No general solution to (2.5) is available although solution for some special cases of $h_a(t)$ have been found[1]. If $\phi_i(t)$ is superimposed over a static frequency offset Δw_i at the input and the loop remains in 'lock' (i.e., the VCO phase $\phi_o(t)$ equals input phase $\phi_i(t)$ within some small margin of error), the linear model can still be applied by reducing the gain of the phase detector from K_d to $K_d \mathrm{Cos}\phi_s$, where ϕ_s is the static phase error due to the input frequency offset. The static error ϕ_s is given by:

$$\frac{\mathrm{Sin}\;\phi_s}{\Delta w_i} = -1/K_o K_d K_a \qquad (2.6)$$

Here, we have assumed the DC gain of $H_a(s) = H_a(0)$ is normalized to unity. Since $\mathrm{Sin}\;\phi_s' \leq 1$, the 'lock-in' or 'hold-in' range defined as the input frequency band over which the Loop is capable of maintaining lock, provided

it is initially locked, is given by:

$$|\Delta w_i|_{\text{Hold-in}} \leq K_o K_d K_a = K \text{ rad/sec.} \qquad (2.7)$$

where K is the total loop gain:

Open & Closed Loop Frequency Response:

From the Linear Model in Fig. 2.3, we have:

$$G(s) = \left.\frac{\phi_o(s)}{\phi_i(s)}\right|_{\text{open loop}} = \frac{KH_a(s)}{s} \qquad (2.8)$$

$$H(s) = \left.\frac{\phi_o(s)}{\phi_i(s)}\right|_{\text{closed loop}} = \frac{KH_a(s)}{s+KH_a(s)} \qquad (2.9)$$

Depending on the choice of $H_a(s)$, we get different types and orders of Loops. By 'type', we mean the number of poles at the origin of the open Loop response and by 'order', the degree of the denominator polynomial of the Closed Loop response.

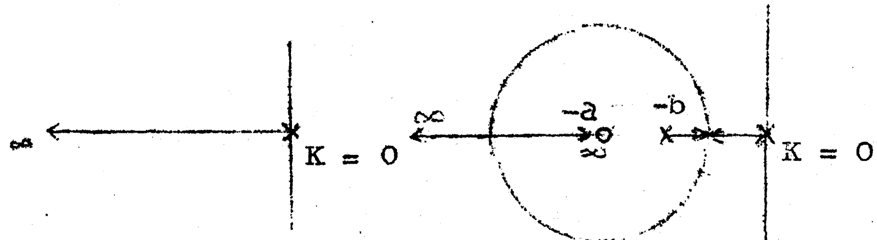
I. First Order Type One Loop:

For this, $H_a(s) = 1$ (2.10)

This gives $H(s) = 1/(1+s/K)$ (2.11)

The root-locus plot of this is shown in Fig. 2.4(a) Here, the pole for the closed loop in on the negative real axis and equals the loop-gain. In this case, we

cannot vary bandwidth and gain independently, since the bandwidth equal to the gain.



(a) First order loop (b) Second Order type one loop

Fig. 2.4

II. Second-Order Type One Loop:

$$\text{For this, } H_a(s) = \frac{(1+s/a)}{(1+s/b)} \tag{2.12}$$

A practical way to realize this is shown in Fig. 2.5:

For this:

$$H(s) = \frac{Kb\ (\frac{s}{a}+1)}{s^2 + Kb\ (\frac{1}{K}+\frac{1}{a})s + Kb} \tag{2.13}$$

Writing this in standard form for 2nd. order system

$$H(s) = \frac{w_o^2\ (1 + s/a_o w_o)}{s^2 + 2\zeta w_o s + w_o^2} \tag{2.14}$$

we have: Loop natural frequency $w_o = \sqrt{Kb}$

Damping factor $\zeta = \frac{1}{2}\left[(b/K)^{\frac{1}{2}} + (Kb)^{\frac{1}{2}}/a\right]$ (2.15b)

$$a_o = a/\sqrt{Kb} \qquad (2.15c)$$

The root locus for this is shown in Fig. 2.4(b).

From the root-locus, we observe that this loop is unconditionally stable. The closed loop bandwidth as determined by w_o and ζ can be chosen independently of the loop gain. These features make the 2nd. order type one loop very attractive in practice. As limiting cases, we get the Ist. order loop it $b/w_o = 1$. As $b \to 0$, we get the second order type two loop. All loops of order greater than two are only conditionally stable. Details about higher order loops can be found in the literature[2]. In the following, we discuss about the design of PLL using 2nd. order type one loops only. Some designs for third and higher order loops have been reported, but the techniques have not been established as yet .

Non-linear Operating Conditions in the PLL:

When the VCO is not in lock, the behaviour of the loop is mathematically complicated. Due to the periodic nature of the phase detector response, no analytical solution is available for the differential equation (2.5) except for the case of first-order loop. For all types of loop, for a

static input frequency the out-of lock condition is characterised by a beat-note at the output of the phase detector (Fig. 2.7).

If the frequency of the beat-note happens to fall close to the bandwidth of the loop filter, the VCO will eventually lock due to the integrating action of the loop filter and the VCO. This range of input frequencies over which the VCO unconditionally locks is called the 'Capture Range' or 'Pull-in Range'. It is usaully slightly greater than twice the loop filter bandwidth except for those which employ ideal integrators in the loop filter. In such cases, the 'Capture Range' equals the 'lock-in range' but the time taken by the VCO to lock may be extremely large. Capture Range is of importance mainly in tracking PLL. In wideband PLDs this is of little consequence.

2.2 Design of PLD for Threshold Extension:

Minimum Loop Noise Bandwidth:

From (2.9) and Fig. 2.3, the output of PLD under linear conditions is given by,

$$C(s) = s\,H(s)\,\phi_i(s)/K_o \qquad (2.16)$$

For a well designed loop, the loop response is constant over the basebandwidth. In such a case (2.16) is equivalent to a conventional discriminator followed by the LPF $H(s)/K_o$. Hence under high CNR conditions, the output SNR of the PLD is identical to the conventional Limiter-Discriminator. In this case, the minimization of output noise implies minimizing the equivalent noise bandwidth of the loop response H(s).

For the second-order, type one loop (2.14), the noise bandwidth is given by:

$$B_n = \frac{1}{2\pi H_o^2} \int_0^{\infty} |H(jw)|^2 dw = \frac{Kb\left(\frac{Kb}{a} + a\right)}{4a\left(\frac{Kb}{a} + b\right)}$$

In most practical cases, $a \ll K$. Then,

$$B_n = \left(\frac{Kb}{a} + a\right)/4 \qquad (2.17)$$

To minimise B_n, we let $\partial B_n/\partial a = 0$, which gives

$$a = (Kb)^{\frac{1}{2}} = w_o \qquad (2.18)$$

This gives one important design equation for the PLL. For this case,

$$B_n = \frac{(Kb)^{\frac{1}{2}}}{2} = w_o/2 \qquad (2.19)$$

The corresponding damping factor is $\zeta = \frac{1}{2}$ (Fig. 2.7)

Threshold Conditions in PLD:

As the input CNR decreases, the probability of the input carrier going through step changes of $\pm 2\pi$ increases. The LPF in the PLD integrates these and the VCO fails to track most of these step changes in phase. It is this suppression of the ThI, which enables the PLD to extend the threshold. The larger the loop bandwidth, the more the fraction of the ThI tracked by the loop. Moreover, when the combined effect of noise and modulation increase the phase error in the loop beyond $\pm \pi/2$ rads., the PLL enters the regenerative region, skips through a cycle of the PD and locks again. This effect contributes a 'click' in the PLD output and is called the 'Loss of Lock Impulse' (LLI). In the design of the PLD, the object is to minimize the total number of impulses due to LLI and ThI. This is achieved if the total mean square phase error due to noise and signal, i.e.

$$\overline{\phi_e^2(t)} = \overline{\phi_{en}^2(t)} + \overline{\phi_{es}^2(t)} \qquad (2.20)$$

is minimized.

The error transfer function for the 2nd. order type one PLD with minimum noise Bandwidth is [(2. 14) and (2. 18) and Fig. (2.3)]

$$\frac{\phi_e(s)}{\phi_i(s)} = 1 - H(s) = \frac{(s/w_o)^2}{1 + (s/w_o)+(s/w_o)^2} \qquad (2.21)$$

For $w < w_o$, the denominator of (2.21) is very close to unity. Then we can rewrite (2.21) approximately as:

$$\frac{\phi_e(s)}{\phi_i(s)} = \frac{s^2}{w_o^2} \qquad (2.22)$$

For FDM signal with baseband spectrum N_m, the input phase modulation spectrum is given by:

$$W_{\phi_i}(w) = \frac{N_m}{w^2} \text{ watts/Hz} \qquad (2.23)$$

The signal induced/mean square phase error is:

$$\overline{\phi_{es}^2(t)} = (1/2\pi) \int_0^{2\pi f_b} W_{\phi_i}(w) \left| \frac{\phi_e}{\phi_i}(jw) \right|^2 dw \qquad (2.24)$$

The noise induced phase error for the minimum B_n PLD(2.19) is:

$$\phi_{en}^2(t) = \frac{2N_o w_o}{2A^2} \qquad (2.25)$$

Adding (2.24) and (2.25), and carrying out the minimization (APPENDIX - A) gives for FDM-FM case,

$$w_o/w_b = 1.61\sigma^{\frac{1}{2}} \qquad (2.26a)$$

$$B_n/f_b = 5.05\sigma^{\frac{1}{2}} \qquad (2.26b)$$

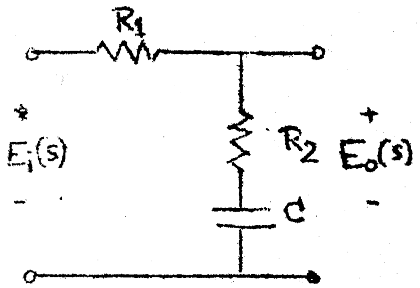
The corresponding threshold value is

$$(CNR)_{AM} = 12.6\sigma^{\frac{1}{2}} \qquad (2.27)$$

Using this value, we can compute SNR in the top channel at threshold. However, these results are optimistic and correct only to a first order (APPENDIX - A). The model assumes that the output noise above threshold in the presence of modulation equals that in the absence of modulation. In practice, the presence of modulation degrades to output NPR by a fraction of 1 dB . The FDM signal is modeled as a baseband Gaussian noise and the modulation by the signal and noise are assumed to be uncorrelated. It is also assumed albeit arbitrarily that the threshold occurs when $\overline{\phi_e^2(t)} = \frac{1}{4}$ rad^2. (this corresponds to input CNR of 4 as referred to PLL noise Bandwidth B_n). It is important to note that PLL performance depends on input CNR referred to loop noise bandwidth B_n and does not depend upon predetection bandpass filter characteristics. The predetection filter is necessary to achieve bandpass limiting or AGC so that the loop gain may be held constant.

Post Detection filtering:

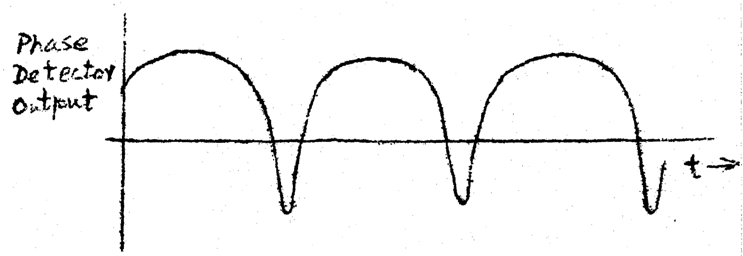
The PLL noise bandwidth B_n, is rather large due to insufficient roll-off of its response (Fig.2.7). In order to



$H_a(s) = (1+R_2CS)/(1+(R_1+R_2)CS)$

$a = 1/R_2C$; $b = 1/(R_1+R_2C)$

Fig - 2.5 : Loop filter



'Out of lock' beat note

Fig. 2.6

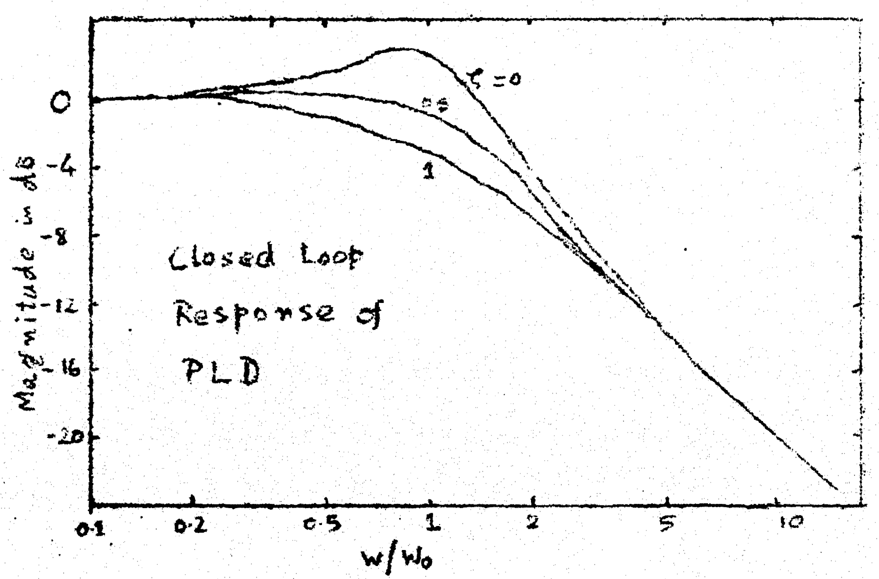
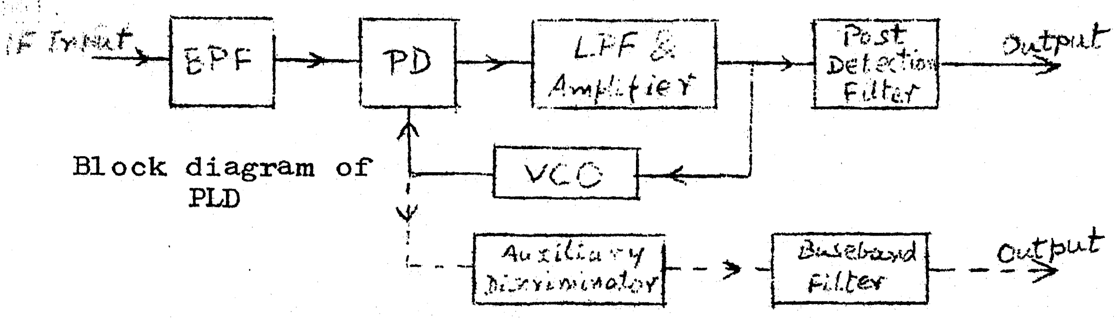


Fig. 2.7

reduce the post-detection noise bandwidth, a rapid roll-off low pass filter is kept after the PLD (Fig. 2. 8). For simple baseband systems (like speech or TV), this results in a noise improvement factor of several dBs. Since the quality of FDM signals depend upon the noise spectrum rather than on total noise power, usually a second-order Butterworth filter is sufficient to compensate for the loop response H(jw).



Block diagram of PLD

Fig. 2.8

Distortions in the PLD:

The main sources of non-linearity in the PLL are the Phase Detector and the transfer characteristics of VCO. The effect of the first one becomes insignificant if the loop gain is kept very high. For the ring-modulator type PD, response can be made more linear if the signal level is made to approach the reference level. Since the nonlinearities in the loop reduce the saturation level NPR at high

CNR, linearizing the PD improves the saturation NPR. Also, the predominant nonlinearity in 'Sine' type PD is the cubic order term, and this is much less severe than square order terms. Non-linearities in the loop generate intermodulation and harmonic distortion products which degrade the maximum NPR obtainable.

Of more importance is the VCO nonlinearity. In many practical wideband schemes varactors are used to construct VCO. Ordinary varactor oscillators exhibit a predominant square law distortion in their voltage to frequency characteristics. The square law distortion introduces largest amount of intermodulation noise in the lower regions of the baseband[2]. If the percentage linearity in terms of 2nd. order distortion is defined as:

$$\Delta = 100\,(a_2/a_1)V_p \qquad (2.28)$$

where V_p is the peak deviation of the VCO in volts, a_1, a_2 are coefficients of the VCO characteristics defined by the infinite series:

$$w(t) = w_c + a_1\,v(t) + a_2(v(t))^2 + a_3(v(t))^3 + \;\ldots\ldots \qquad (2.29)$$

Then the worst case saturation NPR in any given channel is less than

$$NPR = 10^5/2\Delta^2 \qquad (2.30)$$

One way to eliminate 2nd. order and reduce higher order even harmonic distortion is to use a Push-Pull configuration for the VCO. Another method is to use an auxiliary LD (Fig.2.8)

Delay Compensation:

The presence of delay in the loop alters the open loop response to

$$G(s) = H_a(s)e^{-S\tau}/s \qquad (2.31)$$

where τ is the total delay in the loop. With delay, the closed loop response and noise bandwidth B_n become functions of τ. This alters the optimal value of the loop-filter zero 'a'. For the optimum choice of $w_o = 2\ B_n$, if we let

$$a = a_o(Kb)^{\frac{1}{2}} \qquad (2.32)$$

and excess phase shift due to delay at w_o,

$$\phi = w_o = (\phi_b/\pi)\ (B_n/f_b) \qquad (2.33)$$

where $\phi_b = \tau w_b$, is the excess phase shift due to delay at top baseband frequency, then, for the optimal loop with delay, using relation (2.26),

$$\phi = 1.61\,\sigma^{\frac{1}{2}}\ \phi_b \qquad (2.34)$$

The optimal loop noise bandwidth with delay is greater than the one without. This results in a degradation of the threshold. The optimal choice of B_n and 'a' as functions of ϕ are shown in Fig. 2.9. The increase in threshold CNR due to the increase in noise bandwidth may be computed from:

$$(CNR)_{Th} = 10 \log_{10} (B_n'/B_n) \quad dB \qquad (2.35)$$

where B_n' is the noise bandwidth with delay.

Conclusion:

The design of PLD given the IF and baseband specifications consists of (i) finding the optimal loop noise bandwidth and the uncompensated loop filter zero from (2.18), (2.19) and (2.26); (ii) keeping the loop filter pole as low as possible to obtain maximum loop gain; (iii) deriving the worst case linearity tolerable in the VCO from (2.30) and (iv) evaluating the threshold improvement possible from (2.27). The compensation for any appreciable delay present in the loop may be carried out using (2.33) and Fig. 2.9 and the degradation in threshold due to delay is given by (2.35).

References:

1. Intensive Course on Phase Lock Loops - Lecture Notes, Indian Institute of Technology, Kanpur, Dec. 1971.

2. Klapper, J. and Frankle, J.T., "Phase Locked and Frequency Feedback Systems". Academic Press, (1972).

3. International Radio Consultative Committee (CCIR), Recommendation No. 393. Docu. Plenary Assembly, 10th 1963, 4. Int. Telecom. Union, Geneva, 1963.

CHAPTER III

3.1 Design of PLD for a 120 Voice Channel FDM-FM System

In this chapter, the theoretical and practical design and implementation of a PLD having a capacity of 120 voice channels FDM baseband signal, for an Intermediate Frequency of 70 MHz is discussed.

System Design:

The given specifications for a 120 channel FDM-FM link are:

IF Centre frequency	70 MHz
Peak deviation	3.2 MHz
Baseband frequency range	60 to 552 KHz
Peak to meanpower factor for baseband	10 dB

From the specifications we compute the following:

(a) Peak modulation index m_p $= 3.2/0.552 = 5.82$

(b) RMS modulation index σ $= 5.82/(10)^{\frac{1}{2}} = 1.84$

(c) RMS deviation of the carrier Δf_{rms} $= 1.84 \times 0.552 = 1.012$ MHz

(d) IF bandwidth $B_{IF} = 2f_b(1+m_p) = 7.5$ MHz

(e) Threshold CNR:

From relation (2.27) we have,

$$(CNR)_{AM}\Big|_{Th} = 12.6(\sigma)^{\frac{1}{2}} = 17.1 = 12.33 \text{ dB}$$

$$(CNR)_{IF|Th} = \frac{2f_b}{B_{IF}}(CNR)_{AM|Th} = 2.5 = 4 \text{ dB}$$

Comparing with a good limiter Discriminator Threshold of 10 dB, we get threshold extension of 6 dB.

(f) Threshold NPR and Test tone SNR:

From relation (1.15) and substituting for the top channel, $w_{CH} = w_b$,

$$(NPR)_{Th} = 2 \times 1.84^2 \times 17.1 \times 1/0.9$$

$$= 119.7 \text{ or } 21.2 \text{ dB}$$

Assuming a 1 dB threshold degradation,

$$(NPR)_{Th} = 20.2 \text{ dB}$$

This corresponds to a test tone SNR as given by the recommended CCIR[1] formula;

$$(SNR)_{TT} = NPR + BWR - L + 2.5 \text{ dB}$$

where, basebandwidth to channel BW ratio

$$BWR = 10 \log_{10} ((f_b - f_a)/B_c) = 20.9 \text{ dB}$$

Recommended FDM noise loading

$$L = -15 + 10 \log_{10} N \text{ dB} = 5.8 \text{ dB}$$

and N is the number of FDM channels

$$(SNR)_{TT} = 20.2 + 20.9 - 5.8 + 2.5 = 37.8 \text{ dB}$$

PLD Design:

(a) PLL design noise bandwidth:

From the relation,

$$B_n = 5.05\, f_b\, \sigma^{\frac{1}{2}} = 3.78 \text{ MHz}$$

(b) Loop filter zero and pole:

For minimum noise bandwidth, filter zero

$$a = w_n = 2B_n = 7.56 \text{ M rad./sec.}$$

For maximum gain across baseband,

$$b \leq 2\pi \times 60 \text{ KKrad./sec.} = 376 \text{ Krad./sec.}$$

For this, minimum loop gain:

$$K = 4B_n^2/b = 1.51 \times 10^8 \quad \text{sec.}^{-1}$$

(c) VCO Linearity:

For a worst channel NPR of 55 dB at high CNR,

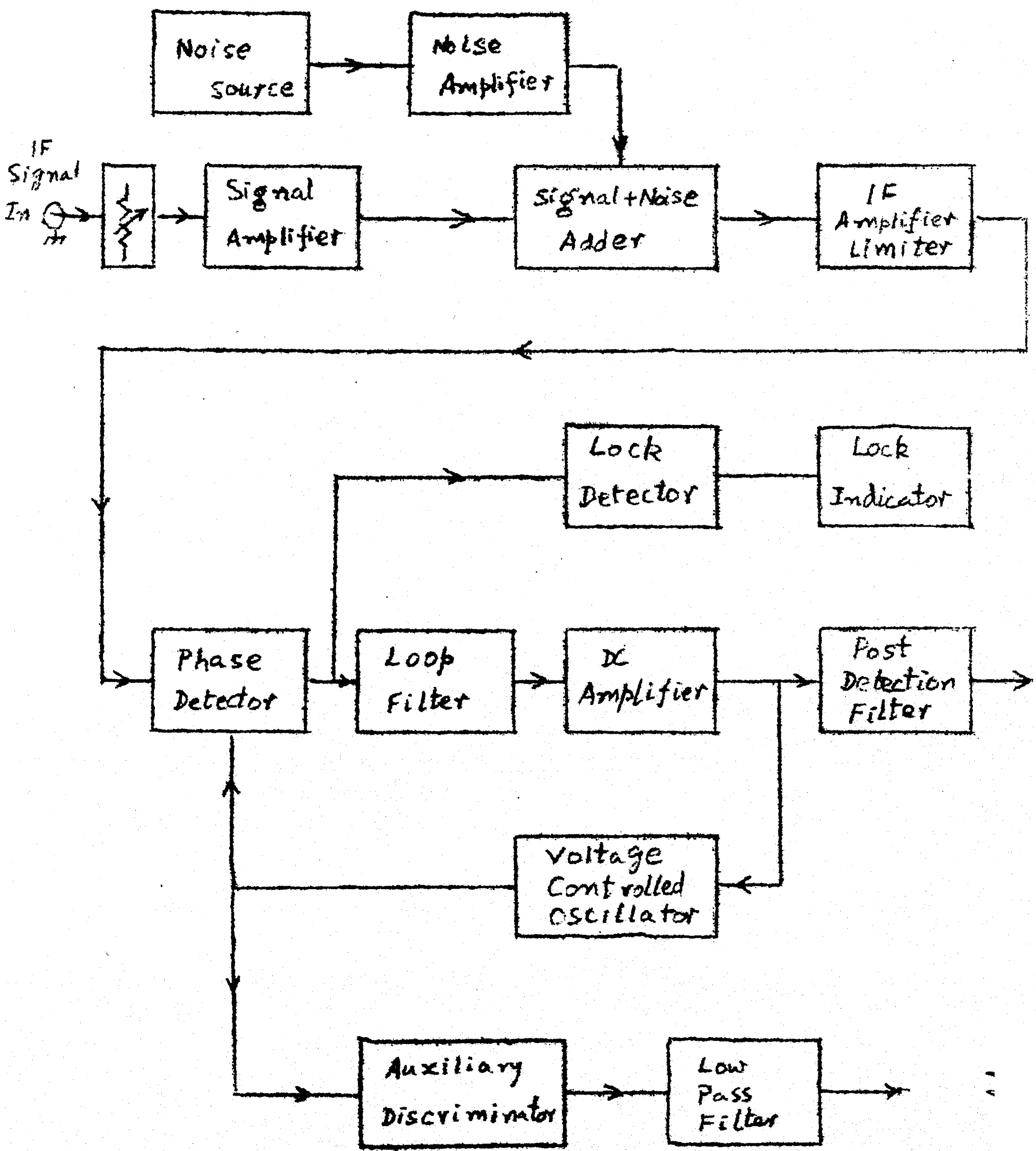
$$\% \text{ Linearity} = (10^5/2\text{NPR})\ \% = 0.4\%$$

3.2 Design of Loop Components:

The overall block diagram of the receiver is shown Fig. 3.1.

Voltage Controlled Oscillator:

The performance of the PLD depends to a great extent on the VCO. It must satisfy most of the following conditions:



Block Diagram of Phase Lock Demodulator

Fig - 3-1

i) Good linearity over the entire IF bandwidth

ii) Minimum delay and very high modulation bandwidth

iii) Resonable stability

iv) Negligible spurious IF output

v) Sufficient power output to drive the phase detector

vi) Constant power output over the entire IF.

The two commonly used wideband VCOs are (i) Reactance tube type, (ii) Varactor Controlled type. With the availability of a variety of varactors the latter one has advantages of simplicity and reliability. One configuration which has been tried as a VCO is shown in Fig. 3.2. This is a common collector Harteley oscillator. The relation between the capacitance and applied voltage for a varactor is given by:

$$C(V_i) = b(V_d + V_i)^{-r} \tag{3.1}$$

where b is a constant of the varactor

V_d is the diffusion voltage of the varactor

V_i is the applied voltage

r is the characteristic exponent of the varactor

r is usually in the range 0.3 to 0.8 for diffused junction varactors and 1 to 2.5 for hyperabrupt junctions. With modulation voltage V_i added on to a bias of E_b volts, the

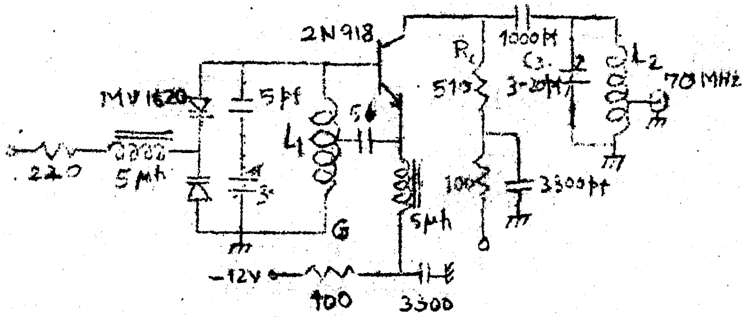
voltage frequency relation is given by the power series[2].

$$w = w_c + a_1 V_i + a_2 V_i^2 + \ldots + a_n V_i^n \qquad (3.2)$$

where, $$a_n = (1/(Lb)^{\frac{1}{2}})\ V_t^{(r/2 - n)} \prod_{k=0}^{n-1} (r/2 - k) \qquad (3.3)$$

L = Associated inductance of the tank (Fig. 3.3)

$V_t = V_d + E_b$ is total varactor bias.



Harteley VCO

Fig. 3.2

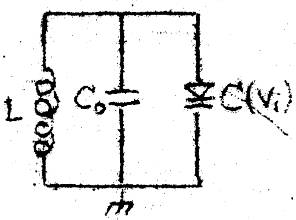


Fig. 3.3

This gives a first order deviation of

$$\Delta w = (r/2V_t)\ W_o V_i \qquad (3.4)$$

The foregoing derivation assumes the simplified model in Fig. 3.3 with total shunt capacitance $C_o \ll C(V_i)$, $Q_{tank} \to \infty$ and negligible stray inductance.

These assumptions are valid in the VHF region. From the relations, it can be shown[2] that

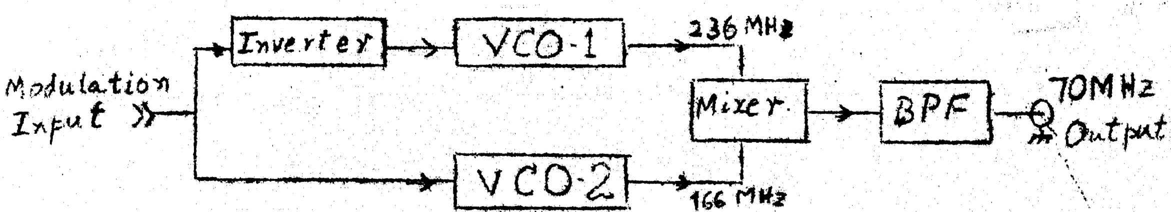
(a) For $r < 1$, the predominant distortion is second order, whereas for $r > 1$ it is possible to compensate for the 2nd. order distortion by series and shunt capacitor combinations.

(b) The sensitivity of the VCO is inversely proportional to the total bias V_t.

In the Harteley oscillator shown in Fig. 3.2 two varactors are used back to back. Since the varactors offer very high impedance to the baseband and the oscillator inductance forms a virtual short for the baseband, the modulation signal cannot modulate the bias of the transistor. This results in minimal amplitude modulation. The common collector configuration offers good isolation between output port and oscillator tank. The bandwidth of the output network consisting of R_c, C_3 and L_2 must be much larger than the peak deviations of the VCO. This is necessary to minimize the delay. The design values for 70 MHz are shown in Fig. 3.2. The V-f_o characteristics obtained is given in Fig. 3.4. One way to reduce the nonlinearity is to use a diode function generator at the modulation port.

Push-Pull VCO:

The second order distortion present in a single VCO using graded junction varactors is effectively cancelled in the Push-Pull configuration (Fig. 3.5).



Block diagram of Push-Pull VCO

Fig. 3.5

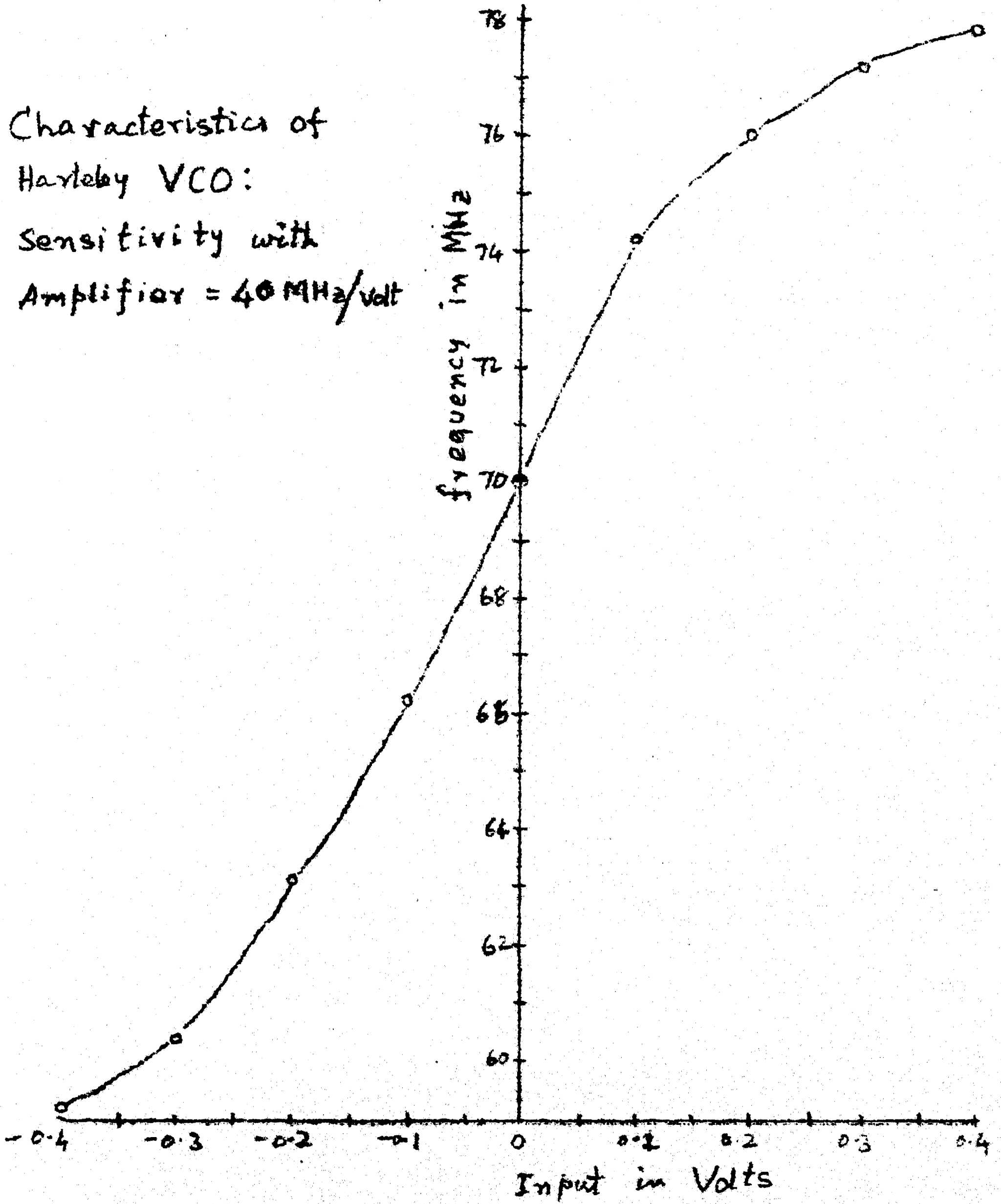

Characteristics of
Harteley VCO:
Sensitivity with
Amplifier = 40 MHz/volt
78
76
74
72
70
68
66
64
62
60
frequency in MHz
-0.4
-0.3
-0.2
-0.1
0
0.1
0.2
0.3
0.4
Input in Volts

Fig. 3.4

All even order distortions are greatly attenuated in the Push-Pull VCO (PPVCO). The PPVCO consists of two VCOs at frequencies f_1 and f_2, with the output taken at their difference ($f_1 - f_2$) by mixing. The two VCOs are driven by the baseband in opposite directions so that the total deviation at IF is the sum of their deviations. Other advantages of the PPVCO are

ii) Since f_1 and f_2 are considerably higher than the IF very good linearity is possible over a wide range of the IF.

iii) For the same reason, very high modulation bandwidth is possible.

iv) As the background noises and the deviations of the two oscillators add arthmetically, a small increase (upto 3 dB) in SNR is possible.

v) Without sacrificing stability, the modulation sensitivity is increased considerably.

For wideband PLD, good linearity, sensitivity and base-bandwidth offered by the PPVCO make it an ideal choice. In order that the mixed IF output of the PPVCO be free from spurious interference products present in a practical mixer, the following choice of oscillator frequencies are recommended[2] for a 70 MHz VCO.

f_1 =	236	254	310	320
f_2 =	166	184	240	250

In our case, the lowest set 236-166 hasbeen chosen.

Design Consideration for the 236 and 166 MHz Oscillators:

The oscillator configuration is the common-base Colpits type (Fig. 3.6). For the varactor Motorola's MV 1620 and for the transistor, 2N 918 were chosen. With a total oscillation capacitance across the tank (C_v = 6 pf at the bias point) of 8 pf, the inductances for the two oscillators are:

$$236 \text{ MHz} : L = 1/(2 \quad x \ 236)^2 \ x \ 10^{12} \ x \ 8 \ x \ 10^{-12} = 60 \text{ nH.}$$

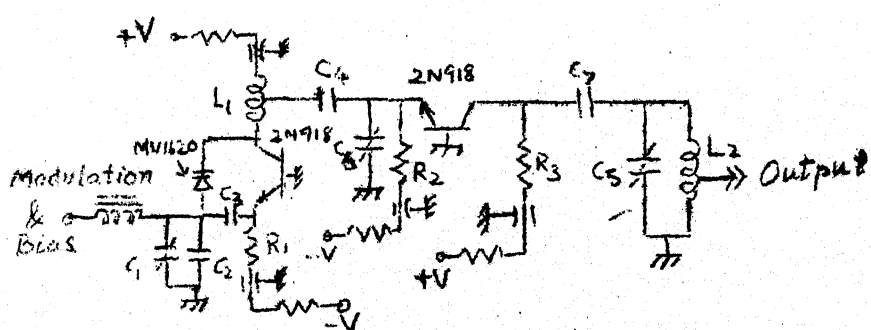
$$166 \text{ MHz} : L = 1/(2 \quad x \ 166)^2 \ x \ 10^{12} \ x \ 8 \ x \ 10^{-12} = 115 \text{ nH.}$$

The dimensions of the inductors are:

60 nH : 2 turns, 1 cm dia, ½" long; Tap at ½ turn from V_+ end

115 nH : 3.5 turns, 1 cm dia 5/8" long; Tap at 0.75 turn from V_+ end

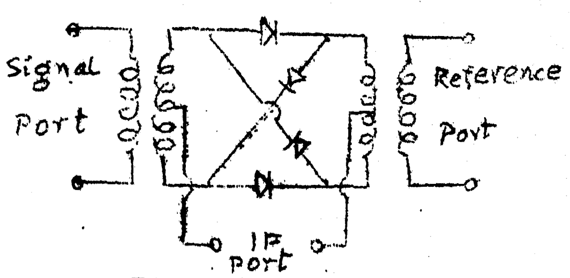
C_3, C_2 and C_1 provide the required amount of feedback. 2N918 with an f > 800 MHz is ideally suited both for the oscillator and the Buffer.



Common base Colpits VCO

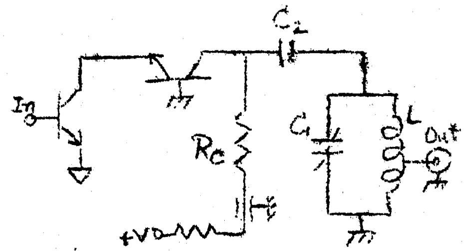
Fig. 3.6

The complete circuit diagram of the PPVCO is shown in Fig. 3.7. Both the oscillators are buffered by two common base amplifiers before driving the mixer. The buffering is necessary to provide more isolation than can be provided by the mixer alone. The mixer chosen is a commercially available ring modulator (ELCOM-DBM 500 PC) consisting of four hot carrier diodes and two wideband ferrite transformers, (Fig. 3.8).



Ring Multiplier

Fig. 3.8



PPVCO Output Stage

Fig. 3.9

The bandpass amplifier following the mixer is a cascode amplifier (Fig. 3.7). The output from the mixer is capacitively coupled to the common emitter transistor. The output network has the following design values:

$f_o = 70$ MHz

Bandwidth = 30 MHz

Loaded Q of tank circuit = 70/30 = 2.33

For the output network shown in Fig. 3.9.

with $R_c = 510$,

$L = R_c/w_oQ = 0.5$ uh.

$C_1 = 1/w_o^2L = 10$ pf.

C_2 = 1500 pf (Coupling)

With L = 8 turns 3/8" dia, for an output impedance of 50 , the tap is at $8 \times (50/510)^{\frac{1}{2}}$ = 2.5 turns from ground.

Baseband input and varactor bias amplifier:

For this a differential amplifier configuration was chosen. This configuration has several advantages over the single input type due to the following reasons:

a) The availability of summing input make the VCO a potentially powerful modem unit. For example, the second input may be used for :

i) AFC injection point

ii) Phase modulation input in a PLL configuration

iii) External tuning point for the VCO

iv) Test input for measuring parameters of a PLD.

b) The paraphase outputs necessary to drive the two oscillators in Push-Pull is natural in a differential amplifier. This obviates the need for an inverter.

Two configurations were tried and both have given adequate performance.

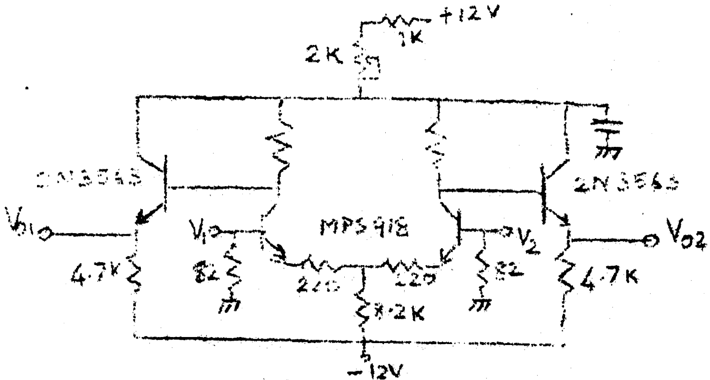
i) Discrete transistor type:

The circuit is shown in Fig. 3.10. In this V_1 and V_2 are the differential inputs. The single ended output gain

is given by $R_{ci}/2(R_e + r_e)$ where r_e is the emitter resistance of the transistor. The variable resistor R_a provides a means to vary the bias on both the oscillator varactors identically. The emitter followers provide low output impedance to drive the capacitive loads of the oscillator inputs over a very wide basebandwidth. This configuration has been used with the Harteley oscillator (Fig. 3.2) in single ended mode. The modulation bandwidth is found to be greater than 15 MHz.

ii) Integrated Circuit type:

For this differential wideband amplifier Fairchild's uA733 was chosen. This hasvery small delay (3.5 ns typical) and a very wide bandwidth (> 90 MHz). Since the output of this floats around + 2 Volts, a biasing network was provided (Fig. 3.11). However, this results in some loss of gain. The input impedance of the VCO is mostly capacitive for the baseband. In order to ensure that the oscillating conditions in the VCO are not disturbed we must provide a high impedance DC path to the varactors. To allow very large base bandwidth. Usually a series RF choke (4 to 10 uh) is used with a resistor (Fig. 3.12) to avoid peaking in the baseband response. The final characteristics of the PPVCO are given in Table 3.2 and Fig. 3.13. The values obtained are:



VCO driver stage

Fig. 3.10

i) Sensitivity K_o = 26 MHz/Volt,

ii) Power output = 5 dBm.

iii) Center frequency = 70 MHz

iv) Static linear range = 70 $\pm$ 10 MHz

v) Modulation Bandwidth 15 MHz

vi) Distortion $<$ 0.5% over $\pm$ 5 MHz.

Phase Detector:

The Phase Detector used in wideband PLD must have the following characteristics:

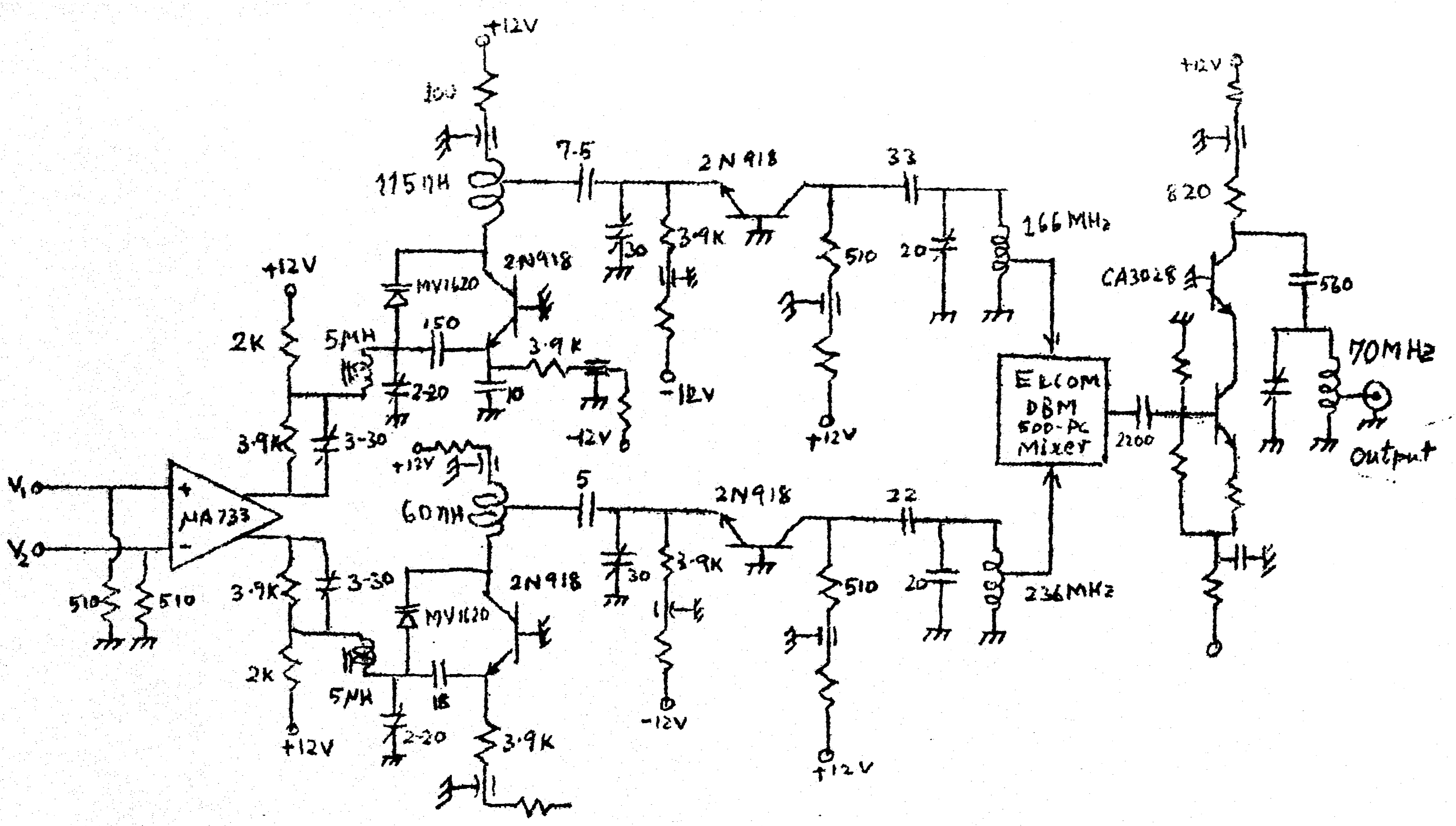
i) Large output bandwidth

ii) Reasonable gain

iii) DC balance and symmetry

iv) Good dynamic range, i.e., the output swing must get doubled for every doubling of the signal input voltage.

All these features are available in high performance commercial ring modulators. For our case, ANIDUS-DBM-1 was chosen. The characteristics of this are given in



Circuit Diagram of Push-Pull Voltage Controlled Oscillator

Fig - 3.7

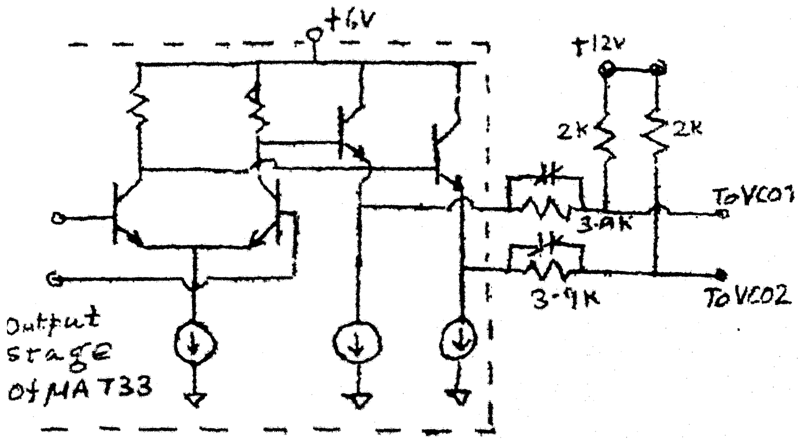


Fig - 3.11

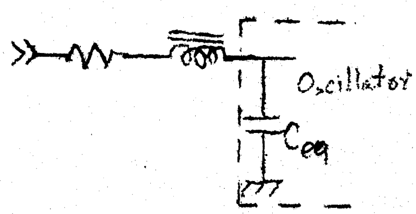
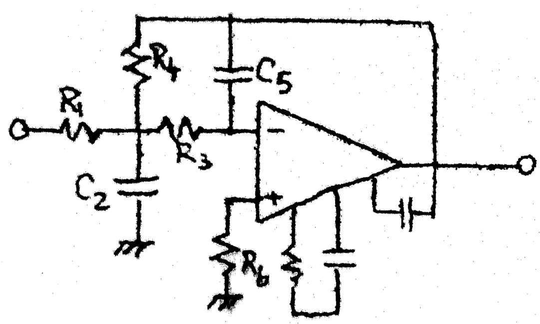


Fig - 3.12



Post Detection Filter

Fig - 3.15

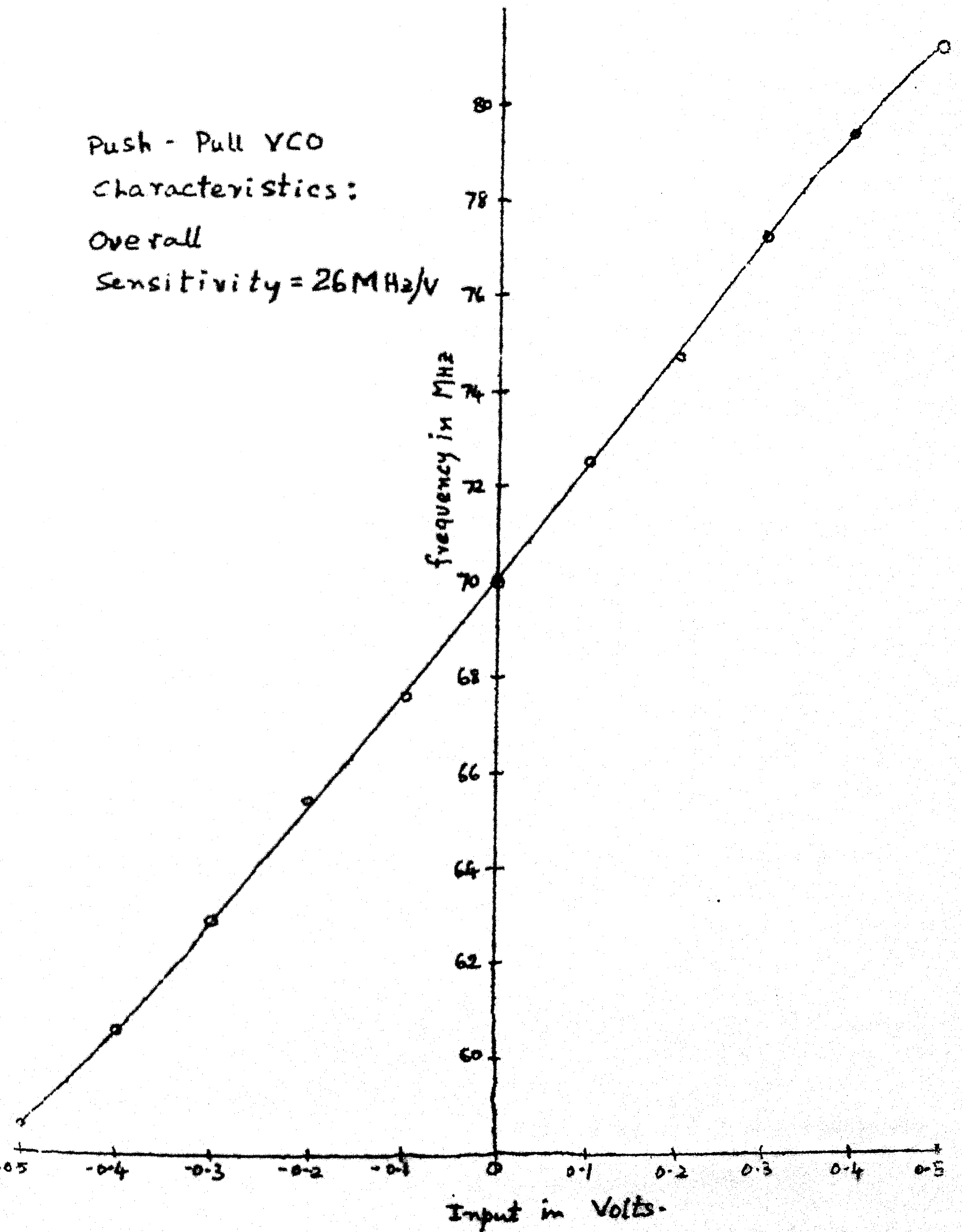

Push - Pull VCO
Characteristics:
Overall
Sensitivity = 26MHz/V
frequency in MHz
80
78
76
74
72
70
68
66
64
62
60
-0.5
-0.4
-0.3
-0.2
-0.1
0
0.1
0.2
0.3
0.4
0.5
Input in Volts.

Fig - 3.13

Table 3.3. In general ring modulators have large dynamic range if the signal port level is considerably lower (about 3 dB or more) than the local oscillator port. When the signal level reaches that of the LO port, the output characteristic becomes triangular instead of sine. For high LO port drive levels (typically 5 dBm) the PD gain is independent of amplitude fluctuations in the VCO.

The obtained values for the phase detector are:

i) LO port drive level = 5 dBm

ii) Signal port drive level = 0 dBm

iii) Sensitivity K_d = 600mv/rad.

Loop filter and Amplifier:

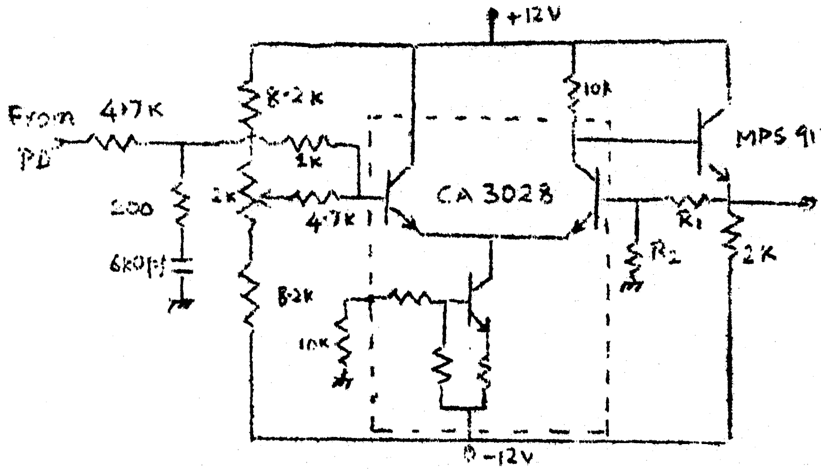
The loop filter is placed immediately after the PD output. From the results obtained in Section 3.1, the filter pole and zero are 376 and 7,560 Krad/sec. respectively. With a choice of C = 680 pf, we get

$$R_2 = 1/7.56 \times 680 \times 10^{-6} = 200 \text{ ohms}$$

$$R_1 = 1/376 \times 680 \times 10^{-9} = 4 \text{ Kohms}$$

The buffer amplifier must be direct coupled, have very wide bandwidth, minimum delay, high input impedance and low output impedance. To achieve this, the circuit shown in

Fig. 3.14 is utilized. It is a negative feedback type non-inverting amplifier, with a gain close to $(1+R_1/R_2)$. The feedback is provided by the resistor divider R_2 and R_1.



Loop Filter and Amplifier

Fig. 3.14

To achieve sufficient input impedance, the differential pair IC is operated at a low level current of 1 ma. To minimize the offset, the resistor R_1 is removed, both bases of the differential pair grounded and R_c is adjusted to give nearly zero volts output. The 2K potentiometer allows small offset adjustments. With R_1 = 1K and R_2 = 220 ohms, this circuit gives an input impedance of 9 Kohms, output impedance less than 10 ohms and bandwidth of 20 MHz.

Post Detection Filter:

For this, a second order Butterworth active filter was designed. With a VCO sensitivity of 26 MHz/Volts, the swing

of the output of the PLD for a $\pm$ 3.2 MHz, deviation is $\pm$ 120 mV_{pp}. For a peak-to-peak swing of $\pm$ 3.6 V, we need a gain of 30. Hence the specifications for the active filter are:

i) DC gain $H_o = 30$

ii) Maximally flat damping factor $a = 0.707$

iii) Corner frequency = 600 KHz = 3.768×10^6 rad./sec.

For the multiple feedback configuration in Fig.3.15

choosing $C = C_2 = C_5 = 56$ pf.

$R_4 = (a/2w_nC)(1+(1+4(H_o-1)/a^2)^{\frac{1}{2}}) = 30$ Kohms

$R_1 = R_4/H_o = 1$Kohms

$R_3 = 1/w_n^2C^2R_4 = 750$ ohms

$R_6 = 1.5$ K

For the Operational Amplifier uA 709 was used.

Lock Detector and Indicator:

In PLDs decrease in input CNR, results in increased cycle slipping. Hence, the average phase error at the output of the phase detector increases with cycle shipping and gives an indication of the nature of lock. To observe this, the simplest way is to amplify the PD output by a wideband amplifier, peak detect the output and use a DC meter to monitor the lock. The wideband amplifier used is similar to the loop amplifier with an extra diode D_1 provided to compensate for the forward drop of the diode D_2(Fig. 3.16).

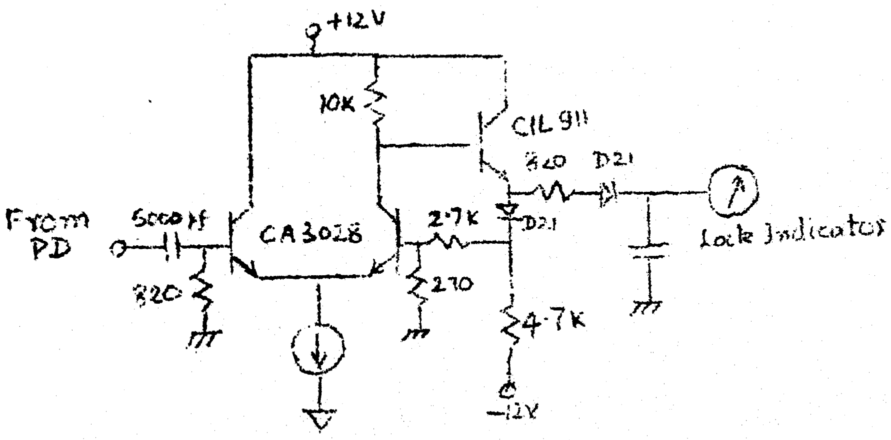


FIG-3.16

3.3 Closing the Loop:

After centering the VCO to 70 MHz, the phase detector is connected to both the VCO and the signal source at its inputs. Then the offset in the loop amplifier is set to zero. The loop is closed by connecting the output of the loop amplifier to the VCO input by a 2K variable resistor which serves as a loop gain control. The output of the PLD will then attain a steady DC level for a static input frequency and lock indicator will show zero output. For an FM input, the output of the post detection filter gives the demodulated output.

3.4 Testing the PLD :

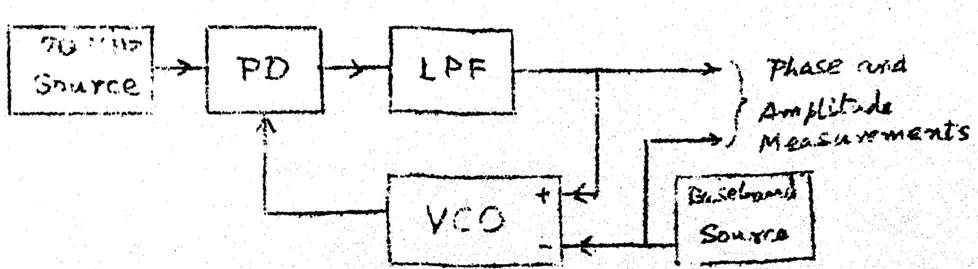
First, the critical components of the PLD, i.e. VCO, PD and the loop amplifier are individually tested. The VCO output is monitored on a CRO, and for varying

input DC conditions it must show a clean sine waveform. Its power output must be constant within the IF bandwidth. The frequency-voltage relation is measured by using a frequency counter.

For testing the Phase Detector, two signal generators are used. The main parameters of interest are (i) output bandwidth as measured by the beat note amplitude against beat note frequency for constant input signal power and (ii) gain. The parameters of the loop amplifier measured are output dynamic range and frequency response.

Closed Loop Test:

Using the set up shown in Fig. 3.17, the closed loop frequency response is measured by varying the frequency of the low frequency signal generator and measuring the amplitude and phase at the output of the loop amplifier. The result obtained is plotted in Fig. 3.18. Any spurious



Test Set-Up

Fig - 3.17

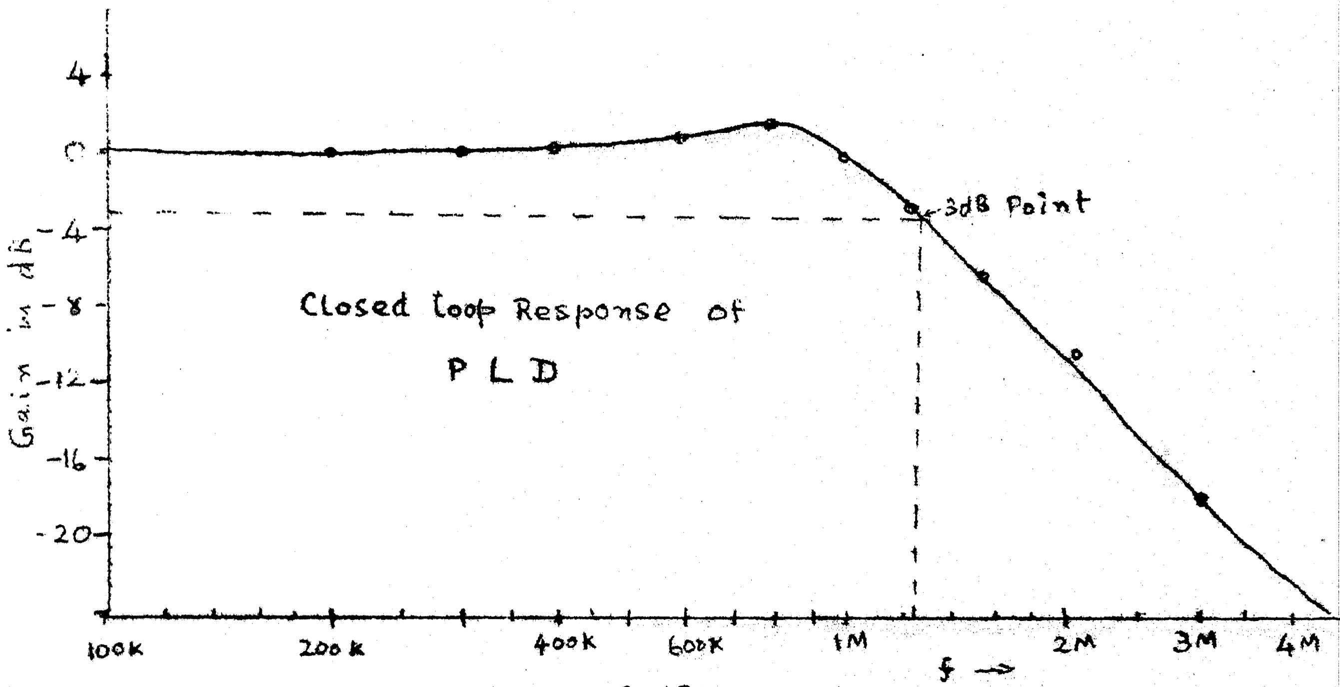


Fig: 3.18

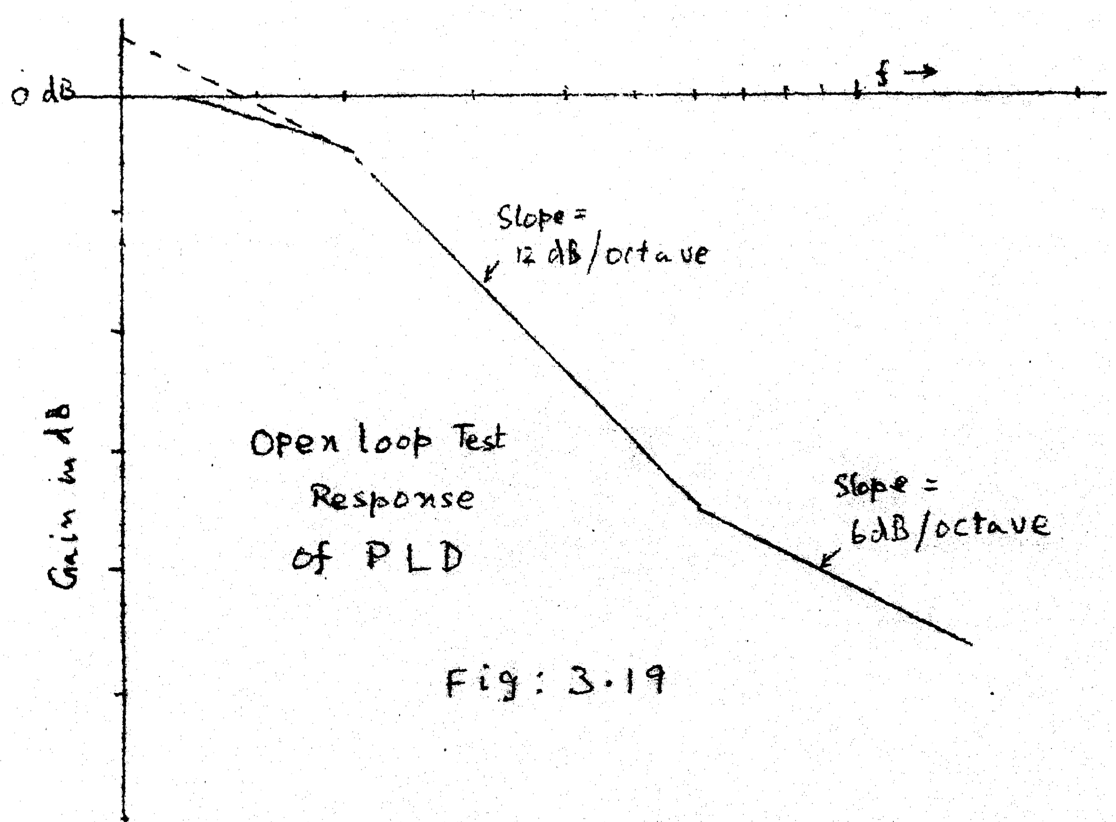


Fig: 3.19

response in the loop can be spotted by measuring the 'open loop response' of the PLD. In the strict sense, the loop can never function under open loop conditions. Yet, if the loop gain is greatly reduced and the closed loop response is measured as before for very low modulation index, the result is very close to open loop response (Fig. 3.19).

References:

1. International Radio Consultative Committee (CCIR), Recommendation No. 393. Docu. Plenary Assembly, 10th, 1963, 4. Int. Telecom. Union, Geneva, 1963.

2. Otala, M., "Distortion and its compensation in a varactor controlled frequency modulator". Proc. IEE, Vol. 117, No.2, Feb. 1970.

3. Otala, M., "The Interference noise in High Quality Push-Pull Frequency Modulators". IEEE Trans. on Com. Tech., Vol. COM-18, No.3, 253-258, June, 1970.

4. Intensive Course on Phase Lock Loops - Lecture Notes, Indian Institute of Technology, Kanpur., Dec. 1971.

Table 3.1

Hartlay VCO.
Voltage - frequency characteristics

Input voltage	f_o MHz
- 0.4	59.28
- 0.3	60.40
- 0.2	63.08
- 0.1	66.21
0.0	70.00
+ 0.1	74.25
+ 0.2	75.97
+ 0.3	77.31
+ 0.4	77.68

Table 3.2

PPVCO Voltage-frequency characteristics:

Input voltage	f_o MHz
-- 0.5	58.57
- 0.4	60.59
- 0.3	62.87
- 0.2	65.28
- 0.1	67.64
- 0.0	70.01
+ 0.1	72.47
+ 0.2	74.82
+ 0.3	77.31
+ 0.4	79.65
+ 0.5	81.49

Table 3.3

Characteristics of Phase Detector: Double Balanced Mixer + ANIDUS DBM - 1

Frequency response	150 KHz to 200 MHz (LO and RF) D.C. to 200 MHz (IF)
Nominal impedance	50 Ohms
Conversion Loss:	8 DB (200 KHz to 175 MHz)
Isolation:	30 DB (50 MHz to 200 MHz)
Minimum LO Power: (refererenced to 50 Ohms)	5 mW (+ 7 dbm) 150 KHz to 100 MHz 10 mW (+ 10dbm) 100 MHz to 200 MHz

APPENDIX - A

Optimum Ratio of Loop Noise Bandwidth to Basebandwidth

The phase error transfer function (2.22) gives,

$$\left| \frac{\phi_e}{\phi_i}(jw) \right| = w^2/w_o^2 \qquad (A-1)$$

From (2.23), (2.24) and (A-1), the signal induced phase error is given by:

$$\begin{aligned} \overline{\phi^2_{es}(t)} &= (2\pi)^2 (N_m/w_o^4) \int_0^{f_b} f^2 df \\ &= (1/3\ w_o^4)\ (2\pi)^2\ N_m f_b^3 \qquad (A-2) \end{aligned}$$

The total mean square phase error is obtained by adding (2.25) and (A-2). Minimization of the total mean square phase error gives the condition:

$$\overline{\phi^2_{en}(t)} = 4\ \overline{\phi^2_{es}(t)} \qquad (A-3)$$

and

$$\overline{\phi^2_e(t)} = (5/4)\ \overline{\phi^2_{en}(t)} \qquad (A-4)$$

Assuming a threshold index mean square phase error of $\frac{1}{4}$ rad^2 and combining (2.19), (2.25) and (A-4), for the minimum noise bandwidth PLL,

$$\overline{\phi^2_{en}(t)} = 2\ B_n N_o/A^2 = 1/5 \qquad (A-5)$$

With as the rms modulation index and $N_m f_b = \sigma^2 w_b^2$, and substituting (A-5) in (A-3) and (A-2) gives the design criterion

$$w_o/w_b = (20/3)^{\frac{1}{4}}\,\sigma^{\frac{1}{2}} = 1.61\,\sigma^{\frac{1}{2}} \qquad \text{(A-6)}$$

$$B_n/f_b = 1.61\pi\sigma^{\frac{1}{2}} = 5.05\,\sigma^{\frac{1}{2}} \qquad \text{(A-7)}$$

The assumption of $\frac{1}{4}$ rad^2 as mean square phase error at threshold is slightly optimistic even for a 'sine' type phase detector. Substituting (A-7) into (A-5) leads to the CNR value at threshold:

$$(\mathrm{CNR})_{AM} = (A^2/2)/2f_bN_o = 12.6\,\sigma^{\frac{1}{2}} \qquad \text{(A-8)}$$

The relations (A-6) - (A-8) form the basic design equations for a second-order type one PLD for low threshold FDM-FM demodulation.

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